261442

(1)

TECHNICAL REPORT

Semiconductor Division

HUGHES

HUGHES AIRCRAFT COMPANY

Newport Beach, California

AUG 17 1961

HIGH FREQUENCY TRANSISTOR STUDY

Contract No. NObsr-85296

United States Navy

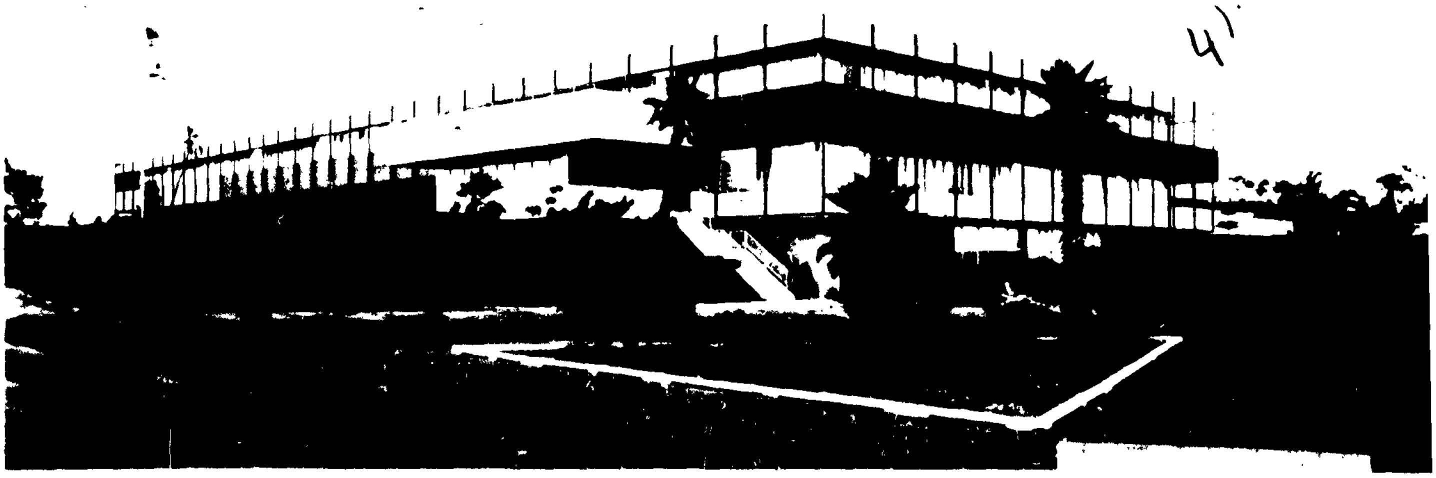
Bureau of Ships

First Interim Engineering Report

10 March through 30 June 1961

61-4-NOX 2

4



REPORT NUMBER 1st Interim Engineering Report

Encl (/) To BuShips Ser 691A1 A5-1047

$3.60

P.40

First Interim Engineering Report

For

HIGH FREQUENCY TRANSISTOR STUDY

This report covers the period 10 March through 30 June 1961

Development Laboratory
HUGHES AIRCRAFT COMPANY
Semiconductor Division
Newport Beach, California

NAVY DEPARTMENT BUREAU OF SHIPS ELECTRONICS DIVISIONS

Contract No. NObsr-85296, Index No. 5B00080302,ST-9349

July 28, 1961

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

Figure 10 Gibbons circuit for transistor operated as a diffusion-delay diode.

Figure 11 Mixer circuit showing V-I characteristics obtained at the emitter with external RF power applied.

ABSTRACT:

A new transistor structure ~~has been~~ designed which will be capable of operation at a fundamental frequency above 1,000MC. Used in conjunction with a coaxial package structure designed for microwave frequencies, the transistor leads can be made sufficiently short to bring the self-resonance of the assembly above the fundamental frequency of oscillation of the transistor. Measurements are reported on present transistors from which the lead inductances of the TO-5 encapsulation are calculated as 3-6nH.

Analysis is given which indicates negative resistance oscillation should be possible for a frequency range where alpha has a phase shift between π and 2π radians. Evaluation of present transistors ~~have~~ verified the presence of this negative resistance oscillation at 1-1.5KMC.

Coaxial mixer circuits ~~have been~~ constructed at Lenkurt Electric Company. Using a ~~Philco~~ coaxial transistor, ~~reduced~~ insertion loss ~~has been~~ achieved at 2KMC.

Prepared by Warren P. Waters

W. P. Waters, Head
Device Development Department

Rainer Zuleeg

R. Zuleeg
Senior Staff Physicist

Approved by Earl L Steele

E. L. Steele, Manager
Development Laboratory

PART I

PURPOSE:

To conduct a research and development program designed to investigate transistor design parameters and circuit requirements which determine and/or influence the parametric mode of transistor operation and to determine feasibility of developing an amplifier, mixer, or harmonic generator operating with high stable gain and low noise at frequencies within the region of 4 to 10KMC. The work shall consist of the following:

TO P. 1

1) Evaluate the present parametric mode transistor design in a suitable microwave mounting without encapsulation for operation in the 4 to 10KMC frequency range.

2) Evaluate available high frequency transistors in circuits designed for operation in the 4 to 10KMC frequency range.

3) Correlate results with theoretical studies and calculate device and circuit interaction.

4) Design and fabricate experimental optimum device structures utilizing advanced techniques for fabrication such as epitaxial growth and oxide masking, and evaluate in circuits for operating in the 4 to 10KMC region.

5) Design circuits for operation in the range of 4 to 10KMC.

6) Evaluate experimental transistors and package structures and determine the capabilities of these transistors for operating in the parametric mode.

7) Determine feasibility of developing an amplifier, mixer, or harmonic generator operating with high stable gain and low noise in the 4 to 10KMC range.

GENERAL FACTUAL DATA

IDENTIFICATION OF TECHNICAL PERSONNEL

	Work Hours
G. Dorosheski, MTS	38.0
C. Fa, MTS	42.0
A. Takacs, MTS	410.0
W. Waters, Department Head	104.0
R. Zuleeg, Senior Staff Physicist	595.0
Lab Technicians	118.0
Research Assistant	51.0
Services	178.0
Other	121.0
Total---	1657.0

REFERENCES

[1] W. W. Gärtner, M. Schüller, "Three-Layer Negative-Resistance and Inductance Semiconductor Diodes", Proc. IRE, 49, 754-767, (April, 1961).

[2] W. Shockley, "Negative Resistance Arising From Transit Time Semiconductor Diodes", BSTJ, Vol. 33, No. 4, (July, 1954).

[3] H. N. Statz, R. C. Pucel, "Negative Resistance in Transistors Based on Transit-Time and Avalance Effects", Proc. IRE, 48, 948-949, (May, 1960).

[4] G. Weinreich, "Transit Time Transistor", Jour. Appl. Phys., 27 No. 9, (September, 1956).

[5] V. W. Vodicka, R. Zuleeg, "Microwave performance of Drift Transistors in the Transit Time Mode", to be published in special issue on Solid State Circuits of the IRE Trans. on Circuit Theory, (December, 1961).

[6] J. F. Gibbons, Stanford University, Stanford, California. (Private communication) to be presented at August 1961 WESCON, San Francisco, "An Analysis of the Modes of Operation of a Simple Transistor Oscillator".

[7] R. V. Pound, "Microwave Mixers", McGraw-Hill, Inc., pp. 92-94, (1948).

DETAIL FACTUAL DATA

INTRODUCTION:

Outstanding mixer performance of commercial Philco transistor T1832 and experimental Hughes mesa transistors has been obtained in a special circuit invented by V. W. Vodicka. Preliminary studies have indicated that the excellent down-conversion properties in certain types of drift transistors are caused by non-linear electrical characteristics with associated negative impedances at the input terminal of a grounded base, self-oscillating transistor. Theoretical studies have been carried out which predict a non-linearity of the total emitter capacitance as a function of current and reveal a "two-three" terminal negative resistance behavior of drift transistors. Since the negative resistance will appear at frequencies that are related to the transit-time of minority carriers across the base region of the transistors, which is itself higher than the cutoff frequency of the device, it was possible to operate the transistor up to the cutoff frequency as an oscillator with good mixing action at harmonics of the fundamental frequency of oscillation. The principle of operation as a mixer and in a transit-time mode are now under detailed investigation and from this study, mixer performance and harmonic power generation above 4,000MC will be explored.

The mixer operation is based on a quasi-parametric, self-pumping effect which led to the designation of "parametric-mode transistor". It is, of course, of importance to show parametric operation in the true sense of the definition and construct a parametric amplifier. This should be indeed possible when using the negative resistance properties together with the non-linear reactances (capacitances) of the transistor [1,2,3,4,5].

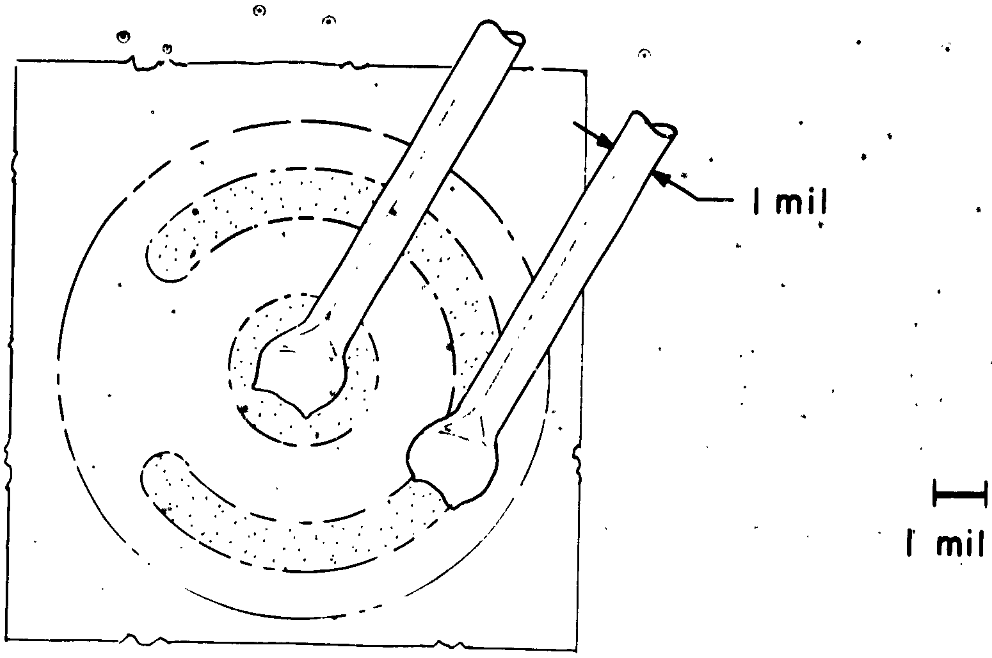
1) Device Design

A developmental Hughes GXG4 Germanium mesa transistor has been used in the exploratory work on mixer performance and harmonic-transit-time mode power generation. This device is packaged in the standard TO-5 encapsulation and microwave operation above 400MC is complicated by the parasitic elements of both transistor and package. Figure 1 is the top view of an assembled transistor with the cross-section of the same unit below it indicating the average dimensions and general material properties. Figure 1a is a photomicrograph of the actual mesa structure, top view. The average electrical parameters of these transistors are:

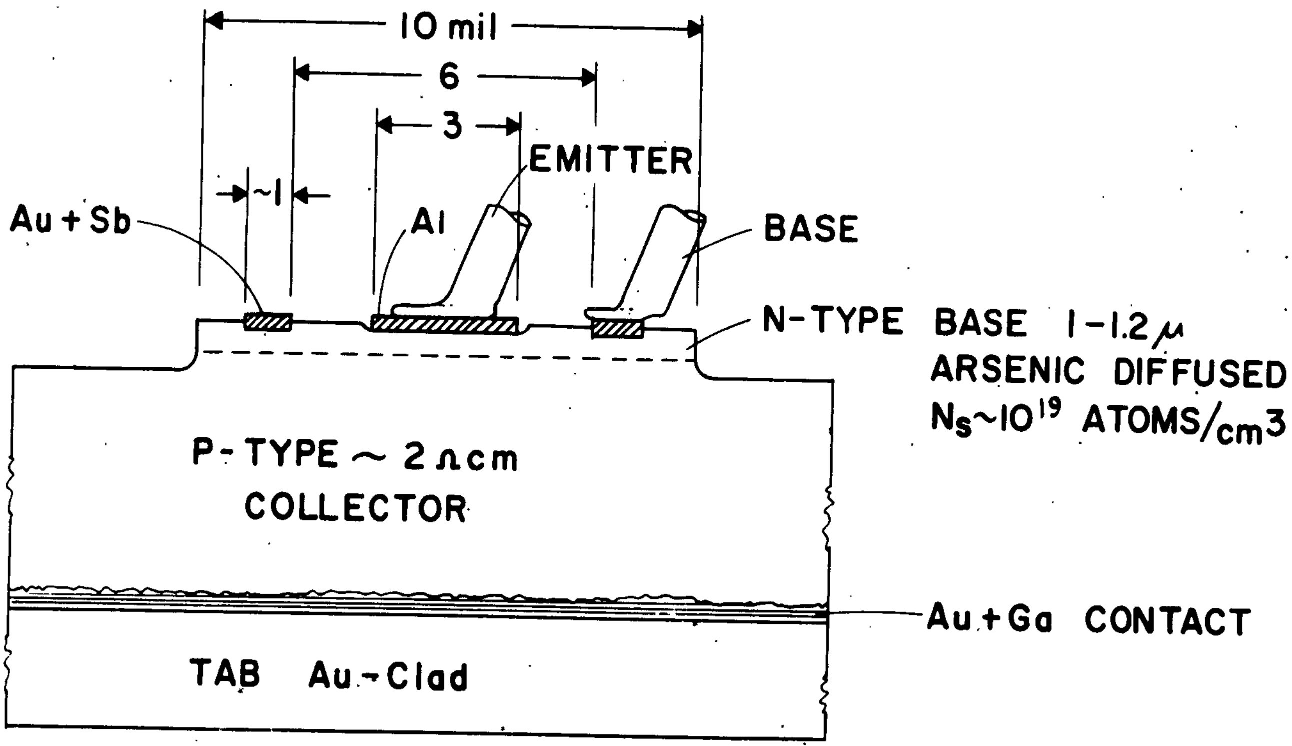
BV_{CB}	=	40-50 volts at 100μA
BV_{EB}	=	.8-3 volts at 100μA
α	=	.92-.985 at V_C = -6V and I_e = 2mA
$f_{c\alpha}$	~	350-400MC at V_c = -10V and I_e = 2mA
f_{max}	~	500-600MC at V_c = -10V and I_e = 2mA
$r_b{}'C_c$	~	40-60μμsec. at V_c = -10V and I_e = 2mA, f = 4MC
C_c	~	4μμf at V_c = -6V and f = 500KC
C_{TE}	~	2-4μμf at V_e = -1V and f = 500KC
R_{CS}	~	30-50 ohms
P_{max} in free air		150mW

From these average measurements and the physical structure of Figure 1a and 1b, a new structure has been designed by extrapolation which should give the following electrical performance:

BV_{CB}	~	40 volts
BV_{EB}	~	.8 volts
α	~	.92
$f_{c\alpha}$	~	800MC
f_{max}	~	1500MC
C_c	~	1-2μμf at V_c = -6V



1 mil
1 mil



10 mil
6
3
~1
EMITTER
Au + Sb
Al
BASE
N-TYPE BASE 1-1.2μ
ARSENIC DIFFUSED
Ns~10^19 ATOMS/cm3
P-TYPE ~ 2 Ωcm
COLLECTOR
Au + Ga CONTACT
TAB Au-Clad

FIG. 1

C_{TE} ~ 1-2$\mu\mu$f at V_E = -.5V

R_{CS} ~ 1-5 ohms

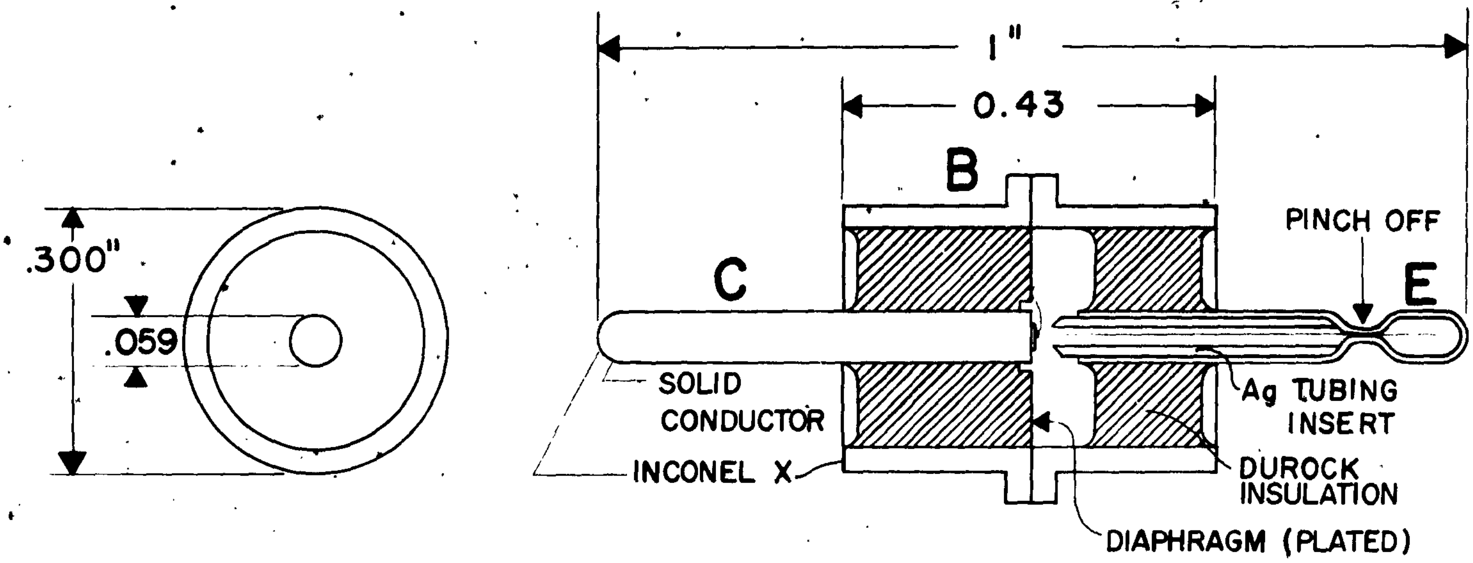
This structure can incorporate epitaxial material and a preliminary run, with the geometry of Figure 1, has indicated the removal of the collector series resistance, R_{CS}, to a few remaining ohms. No transistors of the new geometry have been fabricated. The arrival of the new masks for this geometry is expected for the 1st of August.

2) Package Design

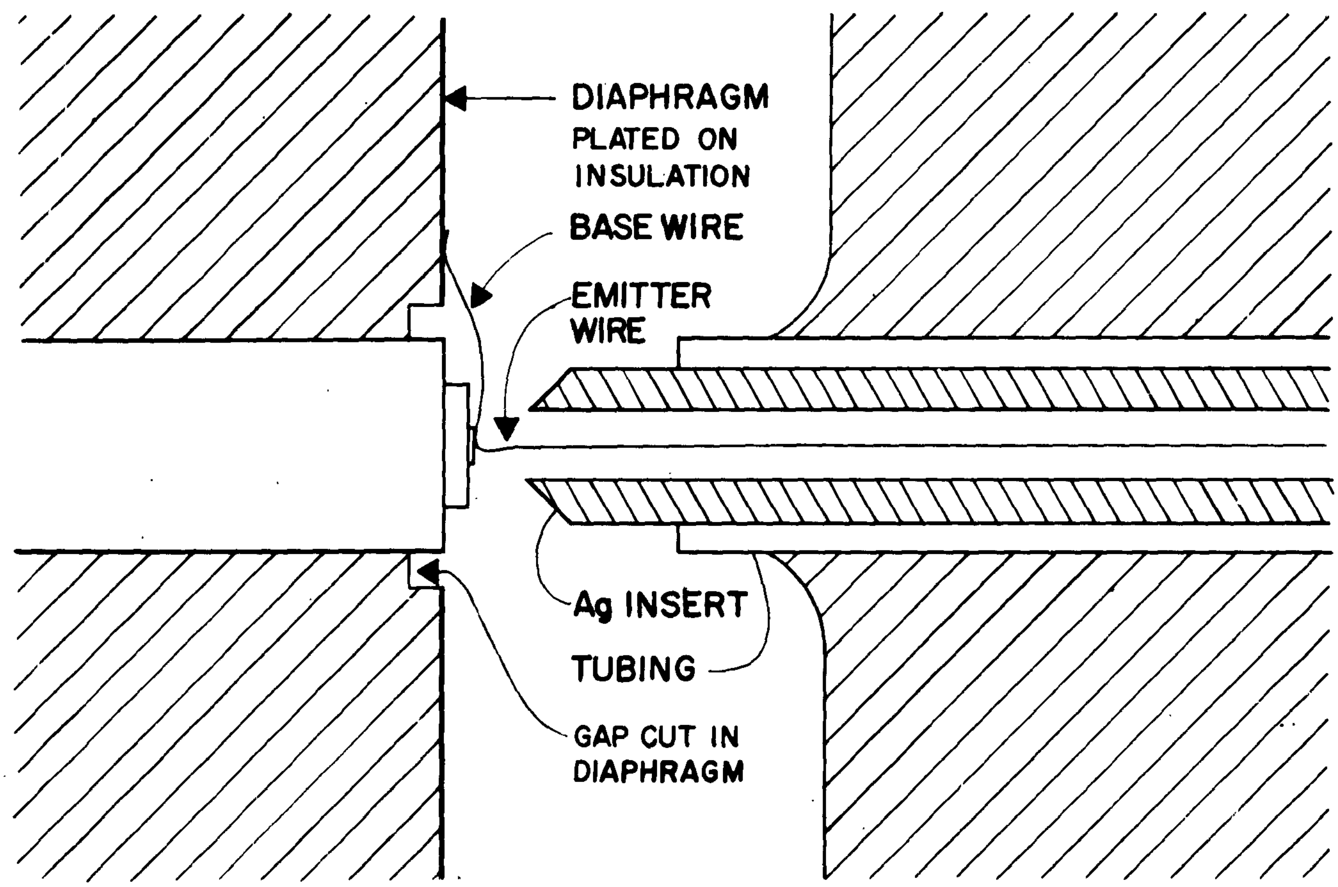
Above 400MC, the TO-5 package introduces extensive parasitic effects of which the interelectrode capacitance is minor, but the lead inductance is a major effect when external tuning is required. The inductances are in the order of 5 - 10 x 10^{-9}H when using the shortest soldered connection without sockets. To overcome these parasitic inductances, two new packaging approaches will be evaluated:

a) Microwave coaxial package
b) MICROSEAL, Hughes dot package

The coaxial package structure is outlined in Figure 2. Outside dimensions close to the Philco coaxial structure are used for socket compatibility. This coaxial structure is designed to present an electrically terminated impedance of about 50 ohms. The diaphragm is plated in a very thin layer across the dielectric to avoid disturbing the impedance. This diaphragm also provides isolation between emitter and collector ends. The epitaxial mesa transistor will be mounted directly onto the center conductor which can be cored with silver to assure good power dissipation properties. The base lead can be bonded to the very thin diaphragm in several places for symmetry. One wire would be only 50-100 mils long and have an inductance of about 10^{-9}H for a 1 mil diameter wire. The emitter will have a single 1 mil wire attached with approximate inductances of 1 - 2 x 10^{-9}H or less,



MICROWAVE COAXIAL TRANSISTOR PACKAGE



DETAIL OF TRANSISTOR LEAD ATTACHMENT

FIG.2

depending on how close the silver tube inserted in the emitter seal can be brought to the mesa. Combined with the new transistor design, it should be possible to operate these transistors in microwave circuits to several KMC.

The Hughes MICROSEAL dot-package, should lend itself to easy incorporation into microwave circuits because of its very small size and it is especially adapted for strip-line circuits. It can be inserted into a coaxial line, or into a cavity arrangment, without disturbing the microwave transmission properties. In addition, the series inductances of emitter and base lead could be made even smaller than in the proposed coaxial package and even better microwave performance could be expected. The design of the MICROSEAL transistor package is outlined in Figure 3 with dimensions given. A special jig has been built to evaluate the package and the transistor in the GR transfer function meter over a frequency range of 200-1500MC.

3) Measurements on present transistors

General Radio transfer-function bridge measurements have been performed on transistors in the TO-5 encapsulation over a frequency range of 200-1500MC. These measurements aided in the determination of the package parasitic elements.

The two important parameters investigated are input impedance with output short circuited and the current-transfer ratio for the common base configuration. A typical measurement of the effective input impedance, $Z_{in(eb)_{eff}}$, is given in Figure 4. The current-transfer ratio has been measured on the active transistor and is called effective alpha, α_{eff}, in contrast to the passive alpha, α_p, which is measured on a transistor with the emitter junction, reverse biased. This establishes a condition in which no minority carrier current flows in the base region and the passive alpha transfer function of the passive network can be determined, i.e., the feedthrough current. The knowledge of the effective alpha and the passive alpha enables one to determine the intrinsic alpha, α_i, of the transistor by vector addition. The intrinsic

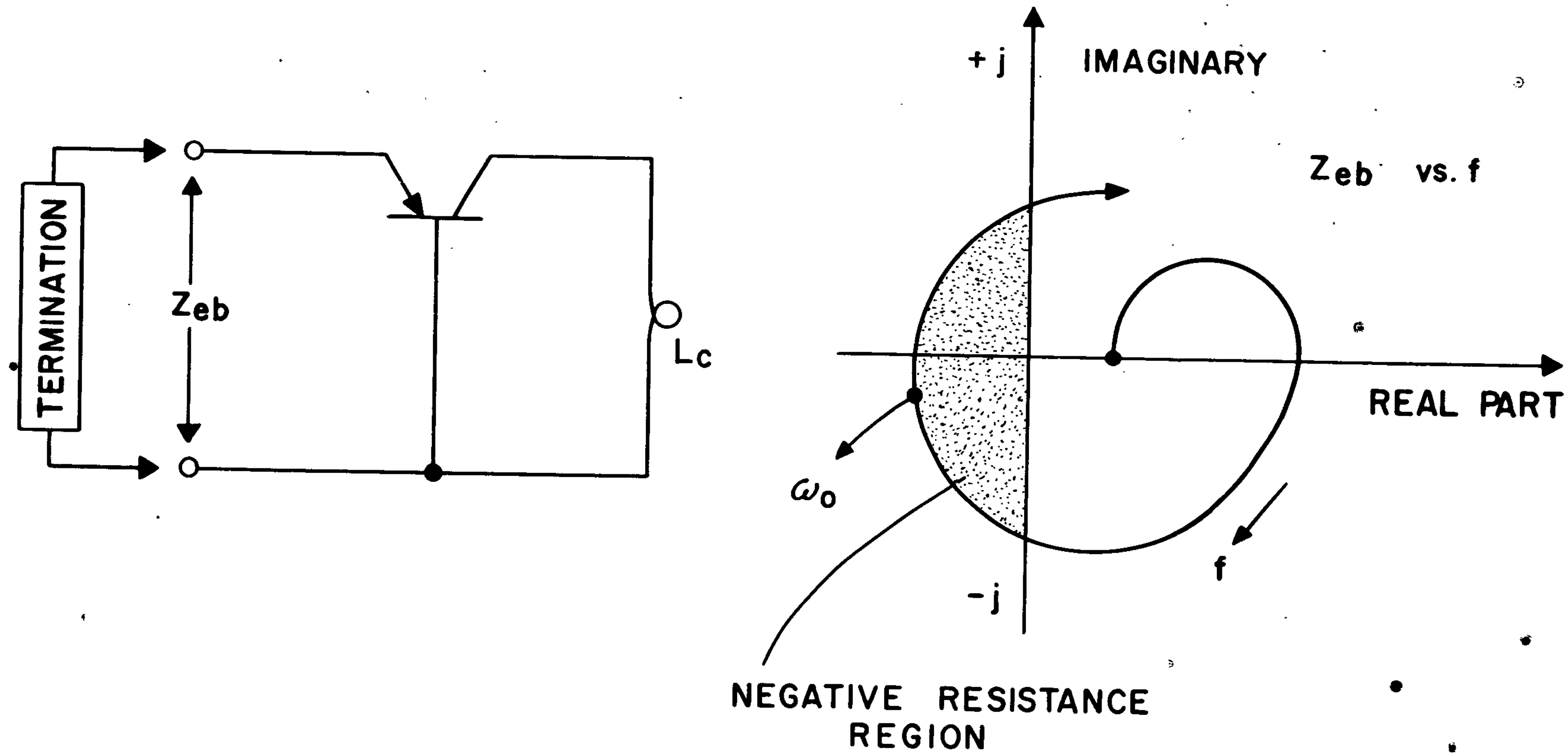

TERMINATION
Zeb
Lc
+ j
IMAGINARY
Zeb vs. f
REAL PART
ω0
f
-j
NEGATIVE RESISTANCE
REGION

FIG. 9

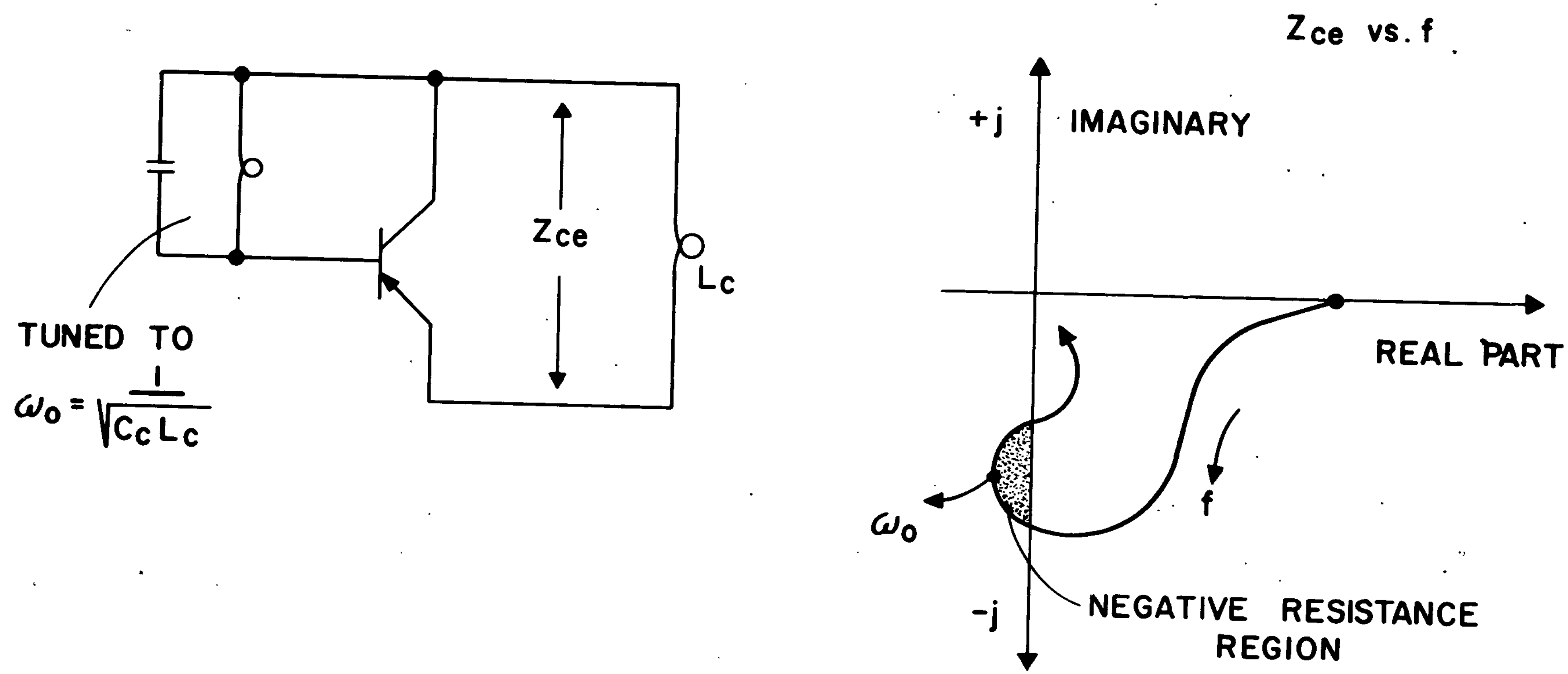
A circuit in Figure 10 was proposed by J. F. Gibbons[6]. In this arrangement, the collector-to-emitter connection is terminated into an inductance, L_c, with the base-to-collector circuit open circuited at the particular frequency so that no RF current is flowing in the base. In this mode, the transistor is operated as a diffusion-delay diode. Oscillations are possible in a frequency range where α has a phase shift of $180° < (\omega) < 360°$. The circuits of Figures 8 and 9 have been constructed with coaxial line elements and the negative resistance oscillation verified. Power outputs into 50 ohms loads at 1000-1500MC of about 1-1.5mW could be obtained. The operation, however, has not yet been made satisfactory since lower frequency oscillation which are not harmonically related to ω_o, are also present and are attributed to the parasitic elements of C_c, L_e, L_b, which cause low frequency resonance around 300-400MC. It is hoped that with the new structure in the coaxial package, clean oscillations in the transit-time mode can be achieved.

5) Circuit studies

To obtain more information about the mixer properties of the emitter-to-base diode in the experimental drift transistors, a circuit arrangement of Figure 11 has been used. It has been found that by external pumping, certain transistors in a given frequency range exhibited the V_{eb} - I_e characteristic of Figure 11c. This could be accomplished by tuning the coaxial line connected between emitter-to-base terminal at a fixed inductance, L_x, in the collector circuit. The region with negative slope, would have a "negative conductance" for the IF frequency if used as a mixer. Similar behavior has been reported in the literature[7] for diodes and mixer operation has been discussed. The expected electrical performance as a mixer would be:

a) Down conversion or mixing with low conversion losses when the conductance approaches zero and possibly conversion with gain if very low or slightly negative conductances are present.

b) Greatly reduced noise figures.



TUNED TO
$\omega_o = \frac{1}{\sqrt{C_c L_c}}$
Z_{ce}
L_c
Z_{ce} vs. f
+j
IMAGINARY
REAL PART
f
ω_o
-j
NEGATIVE RESISTANCE REGION

FIG. 10

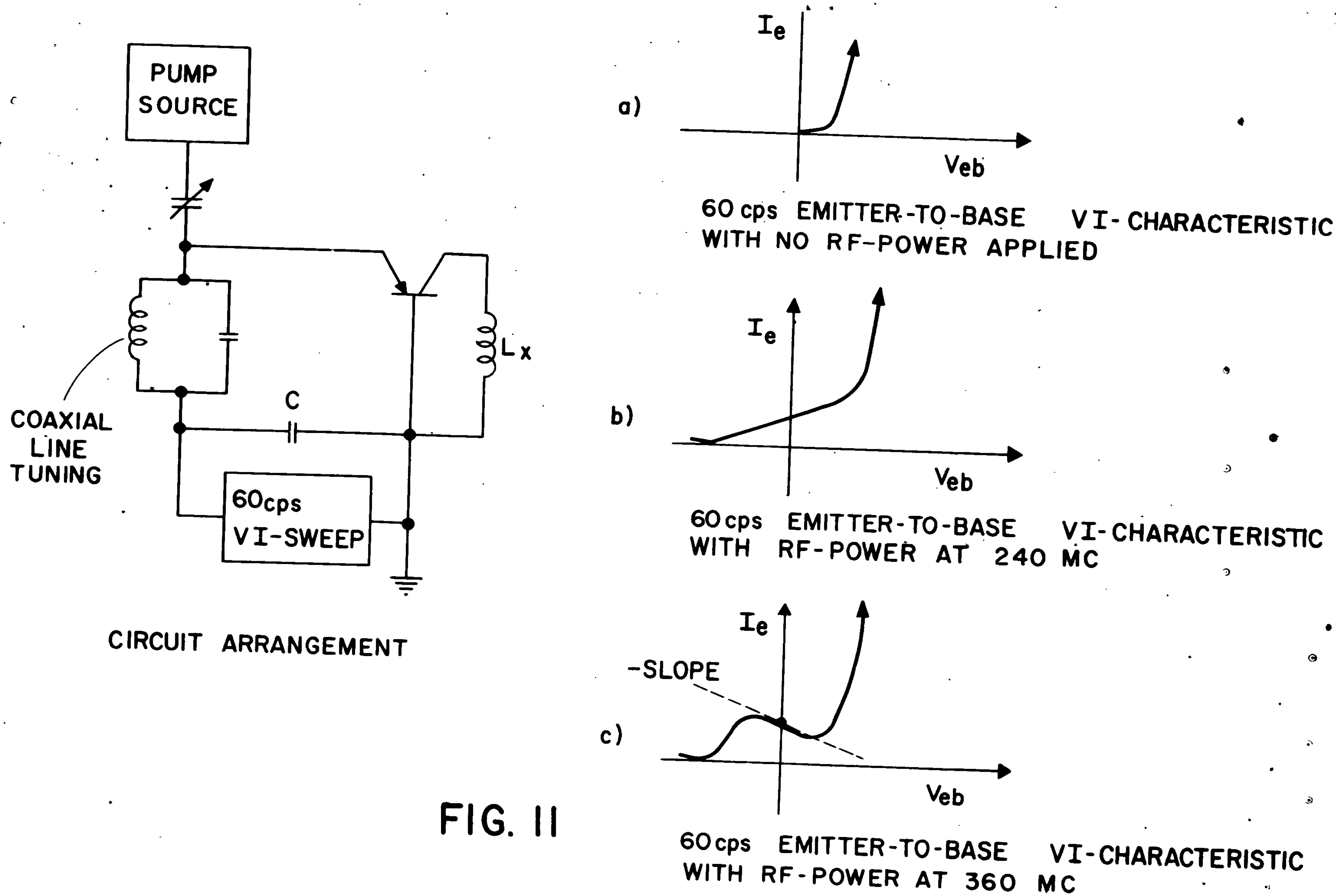

PUMP SOURCE
COAXIAL LINE TUNING
C
60cps VI-SWEEP
Lx
CIRCUIT ARRANGEMENT
a)
Ie
Veb
60 cps EMITTER-TO-BASE VI-CHARACTERISTIC
WITH NO RF-POWER APPLIED
b)
Ie
Veb
60 cps EMITTER-TO-BASE VI-CHARACTERISTIC
WITH RF-POWER AT 240 MC
c)
Ie
-SLOPE
Veb
60 cps EMITTER-TO-BASE VI-CHARACTERISTIC
WITH RF-POWER AT 360 MC

FIG. 11

CONCLUSIONS

Theoretical analysis and measurements which have been made indicate the deleterious effect of the present transistor and package design. Before theoretical studies can be extended, it is essential that experimental transistor measurements be made as a basis for further analysis.

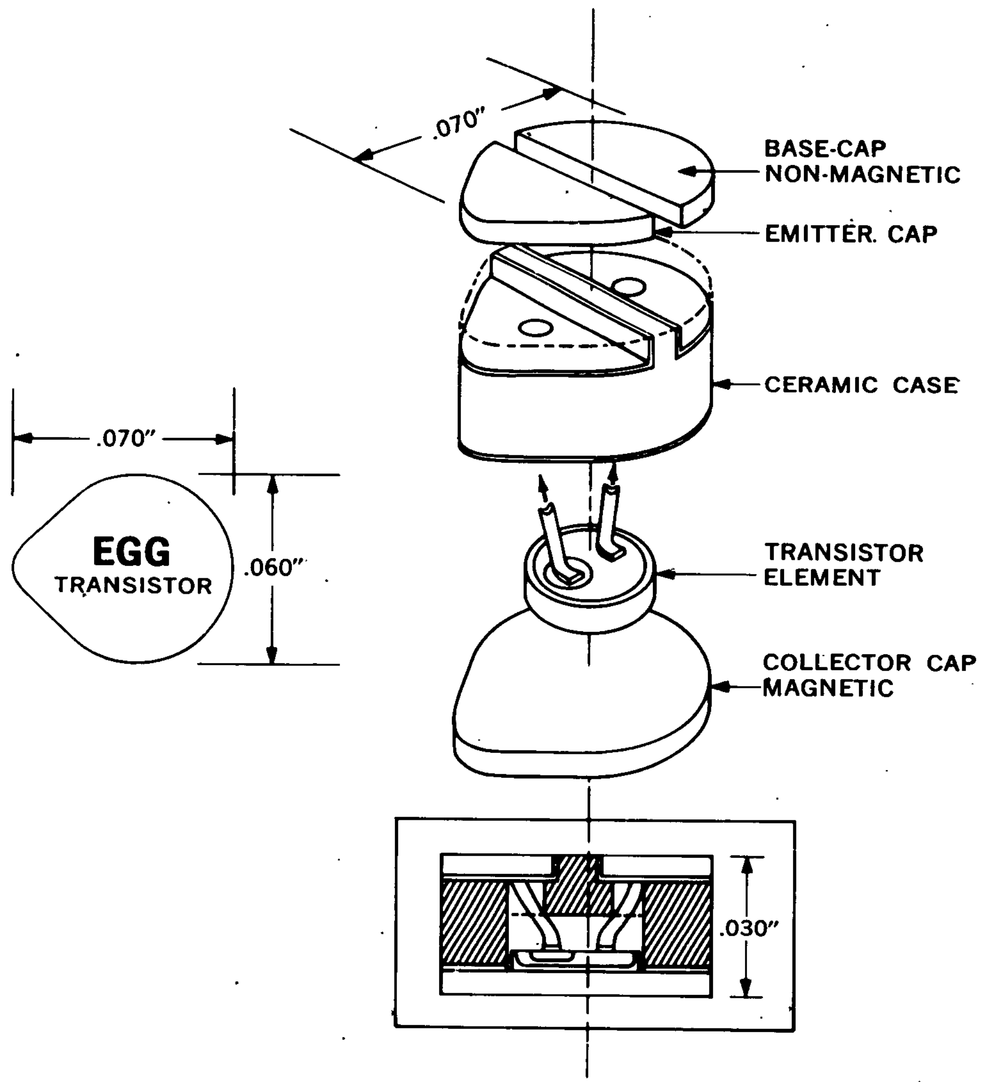
A transistor structure has been designed that is capable of fundamental frequencies of 1500MC from which harmonic operation can be reached above 4KMC. However, it will be necessary to fabricate such transistors in a suitable microwave package to achieve these frequencies in any microwave circuit. The high frequency transistor design requires the use of epitaxial active regions for the mesa fabrication and smaller masks for the optimum structure. The new design must be incorporated in a coaxial, microwave structure which can be used in conjunction with standard microwave equipment and which will minimize lead inductance that now produces parasitic resonance below the desired frequency of oscillation. With the optimum device structure in the coaxial package design, it should be possible to use the negative resistance effects studied in the last quarter to obtain oscillation in the 4KMC range.

PART II

PROGRAM FOR NEXT INTERVAL

Fabrication will be made of transistors with the optimum device structure. The masks for the smaller geometry will be evaluated. It is expected that suitable epitaxial growth with resistivity from 10-20 ohm-cm will be available from another development laboratory program in progress. Six weeks delivery time will be required for experimental microwave package parts which should permit some evaluation before the end of the next period. Measurements will be made, however, of the Hughes MICROSEAL package by the same techniques described in this report as used for the standard TO-5 package. If these measurements prove quite favorable, attempts will be made to encapsulate the optimized device structure into this package. If suitable material can be obtained and time permits, attempts will be made to fabricate a GaAs transistor suitable for operating in the parametric or transit-time modes.

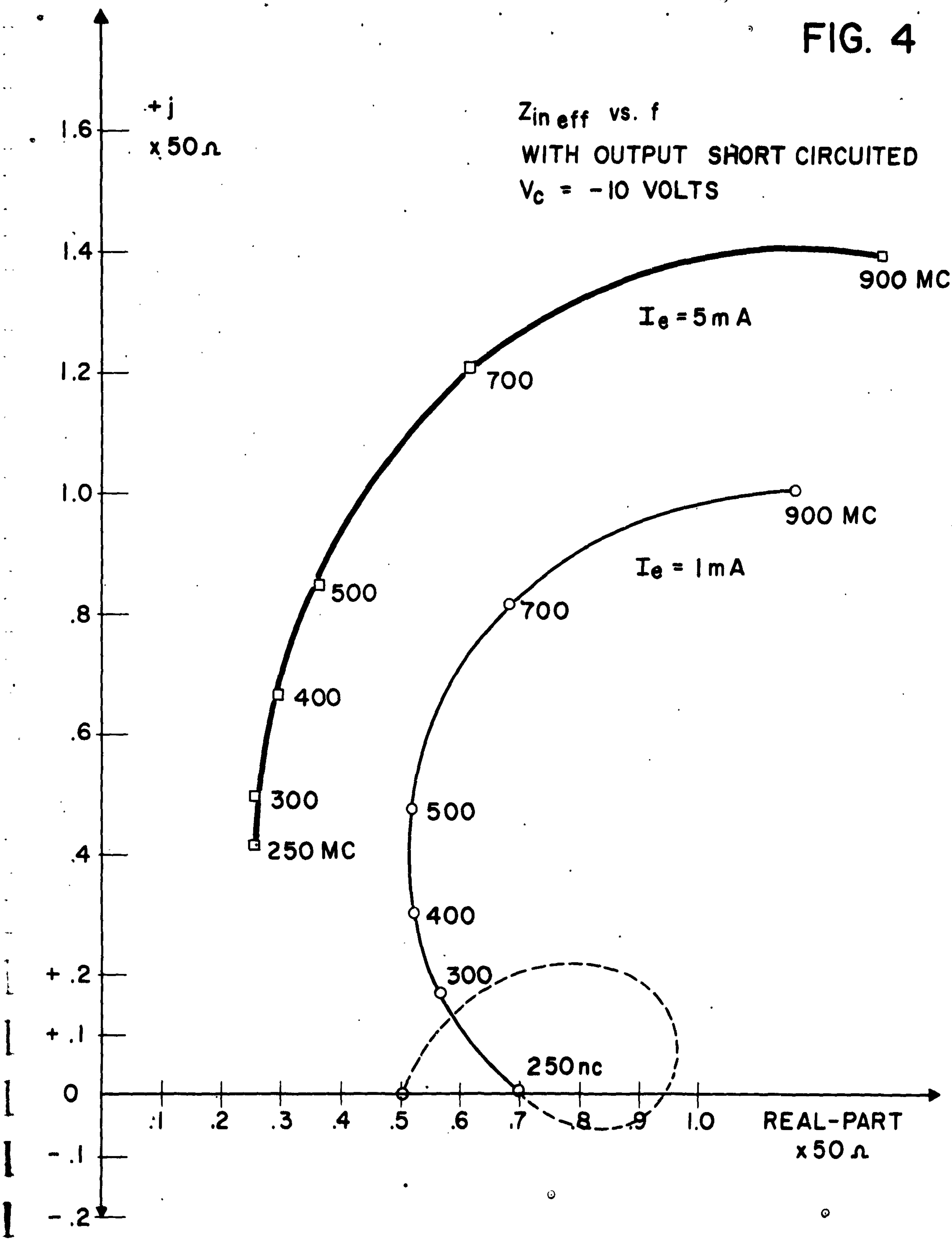
Complete evaluation will be made of the new transistor structure, of the microwave package, and of the effect of the package and transistor combined. If sample GaAs transistors can be made available from the Bureau of Ships Program at Texas Instruments, they will be evaluated for parametric mode transit-time mode operation. As soon as suitable transistors can be fabricated, studies will be made of the negative resistance effects for use in mixer action. Transistors encapsulated in the new package will be forwarded to Lenkurt Electric Company for evaluation in their coaxial mixer cavities.



HUGHES MICROSEAL PACKAGE

FIG. 3

FIG. 4



+j
x 50 Ω
1.6
1.4
1.2
1.0
.8
.6
.4
+ .2
+ .1
0
- .1
- .2
Zin eff vs. f
WITH OUTPUT SHORT CIRCUITED
Vc = -10 VOLTS
900 MC
Ie = 5 mA
700
500
400
300
250 MC
900 MC
Ie = 1 mA
700
500
400
300
250 nc
.1
.2
.3
.4
.5
.6
.7
.8
.9
1.0
REAL-PART
x 50 Ω

alpha is only concerned with the current transfer ratio in the base region of the transistor neglecting external elements. (See Figure 5) Figure 6 gives the effective alpha, α_{eff}, of a typical transistor under operational bias conditions. It has a marked departure from the theoretical behavior of a drift transistor which is indicated by the dashed curve. This deviation from the ideal transistor characteristics is, of course, attributable to the external parasitic circuit elements as given in Figure 5.

The passive alpha, α_p, is given in Figure 7 for the emitter junction with -0.5 volts applied in reverse biased condition. Knowing r_b' and R_{CS} and C_c from separate measurements, it was possible to determine the inductances L_b and L_c. We obtained the following inductance values:

$$L_c = 3.1 \times 10^{-9} H$$

$$L_b = 2.6 \times 10^{-9} H$$

using transistor parameters

$$C_c = 3.2 \times 10^{-12} F$$

$$r_b' = 5 \text{ ohms}$$

$$R_{CS} = 15 \text{ ohms.}$$

A theoretical curve for the passive alpha, transfer function for a transistor with the above values has been plotted as a solid line in Figure 7 from the following relation:

$$\alpha_p = \frac{j\omega\, r_b' C_c - \omega^2 C_c L_c}{1 + j\omega\,(r_b' + R_{CS})\, C_c - \omega^2 C_c\,(L_c + L_b)} \qquad (1)$$

General agreement with the actual measurement has been obtained. Since the effective alpha is represented by

$$\alpha_{eff} = \frac{\alpha_i + j\omega\, r_b' C_c - \omega^2 C_c L_c}{1 + j\omega\,(r_b' + R_{CS})\, C_c - \omega^2 C_c\,(L_c + L_b)} \qquad (2)$$

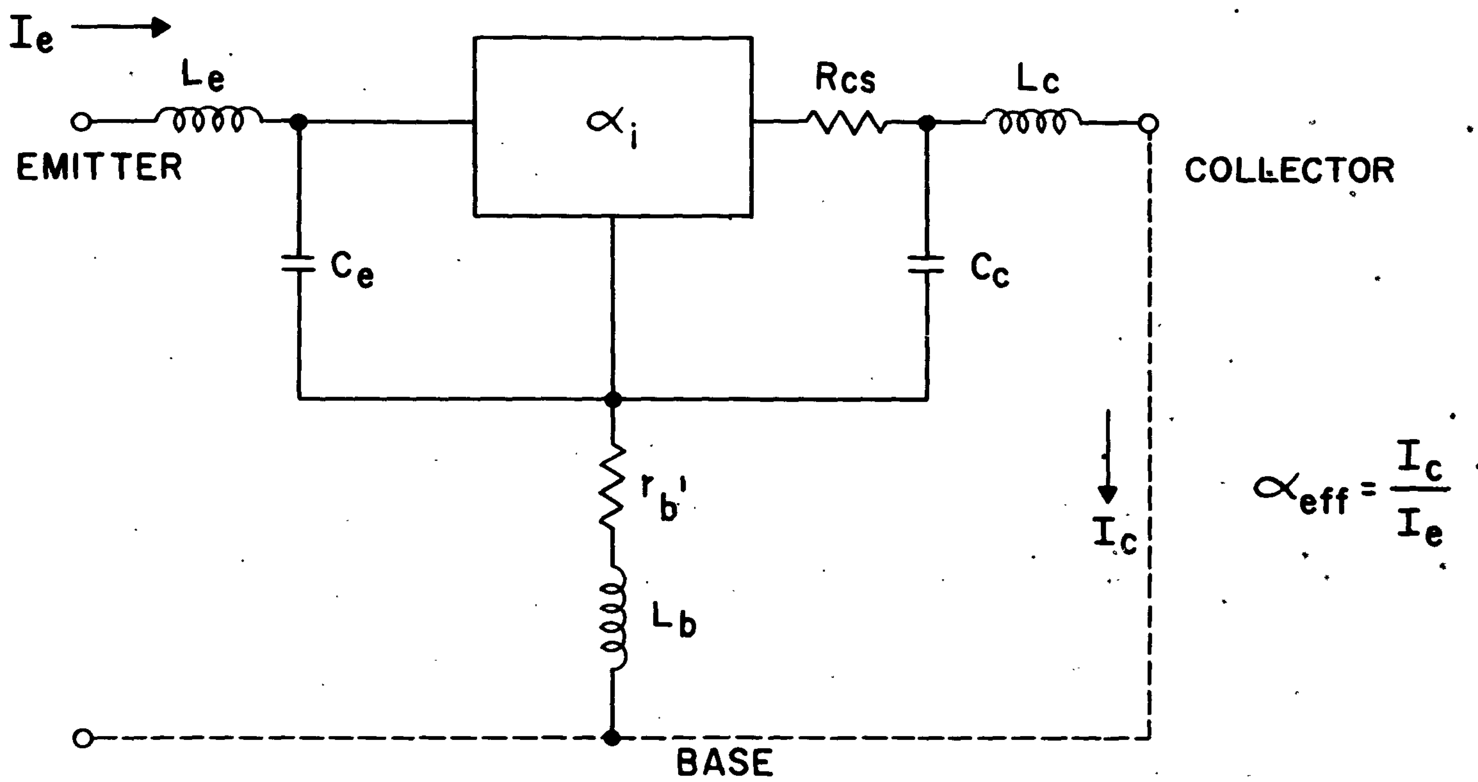
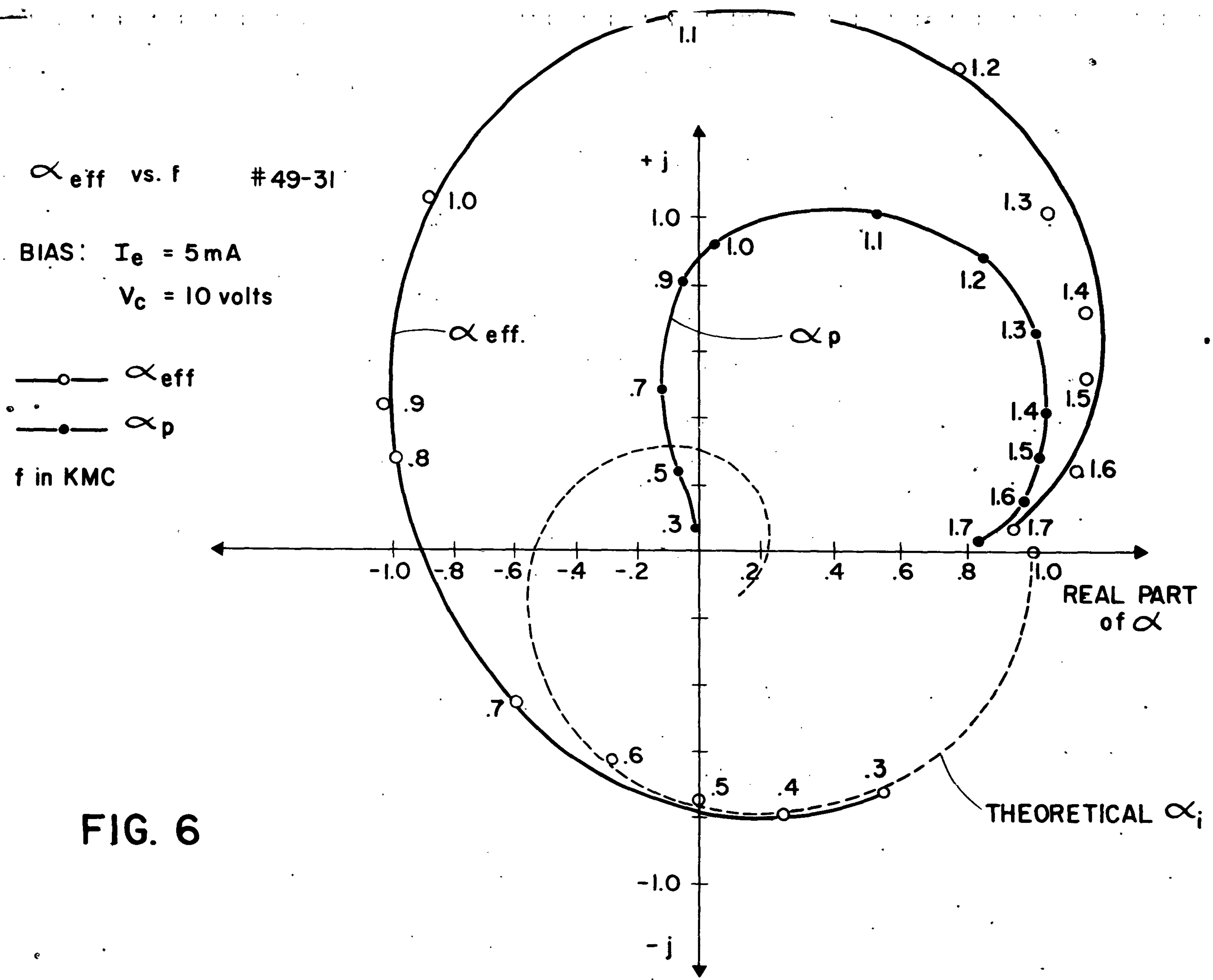

I_e
L_e
EMITTER
α_i
R_{cs}
L_c
COLLECTOR
C_e
C_c
r_b'
L_b
I_c
BASE
$\alpha_{eff} = \frac{I_c}{I_e}$

FIG. 5



α_{eff} vs. f
49-31
BIAS: I_e = 5 mA
V_c = 10 volts
α_{eff}
α_p
f in KMC
α eff.
α p
+ j
- j
REAL PART of α
THEORETICAL α_i

FIG. 6

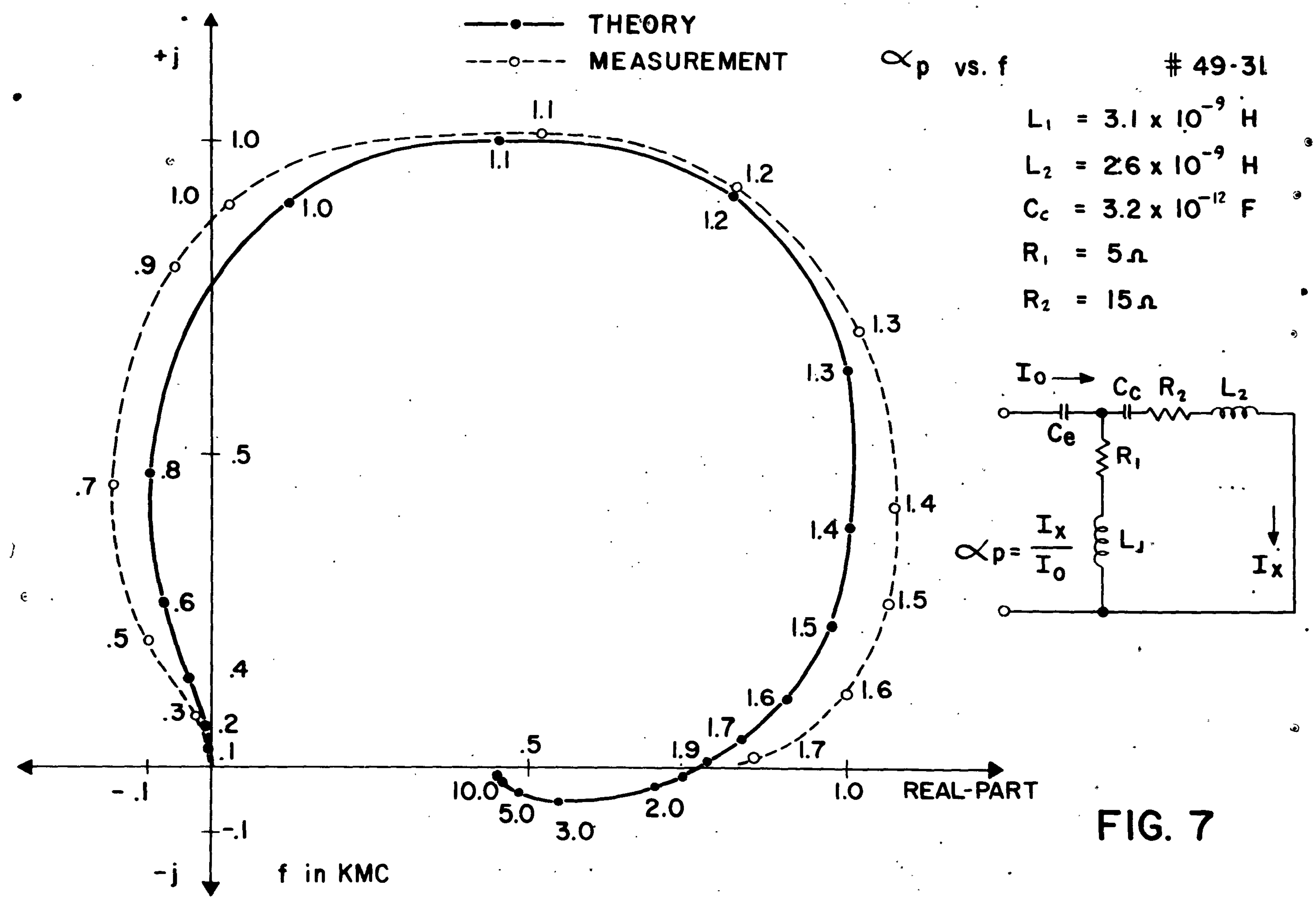

THEORY
MEASUREMENT
α_p vs. f
49-31
$L_1 = 3.1 \times 10^{-9}$ H
$L_2 = 2.6 \times 10^{-9}$ H
$C_c = 3.2 \times 10^{-12}$ F
$R_1 = 5\,\Omega$
$R_2 = 15\,\Omega$
I_0
C_c R_2 L_2
C_e
R_1
L_J
I_x
$\alpha_p = \frac{I_x}{I_0}$
+j
-j
REAL-PART
f in KMC

FIG. 7

it is possible, by suitable analysis in the complex domain, to reconstruct the intrinsic alpha, α_i, and compare it with a theoretical expression

$$\alpha_i = \alpha_o \frac{\exp(-jm\,\omega/\omega_c)}{1 + j\,\omega/\omega_c} \qquad (3)$$

The corrected α_{eff} to yield α_i has been plotted in Figure 8 together with the curve of equation (3) for values of fc = 600MC and m = 1.1. m is the drift field factor and gives the excess phase over 45° at the 3db cutoff frequency point. m = 1.1 corresponds to a drift field of 8KT/q.

The remaining parameter, L_e could be determined from the measurements of Figure 4, when the assumption was made that a series inductance is merely added to the input impedance,

$$Z_{in_{eff}} = Z_{in_i} + j\omega L_e \qquad (4)$$

An estimate of this parameter by curve fitting led to a value of 4-6 x 10^{-9}H.

4) Circuit evaluation

The unilateral gain function, U, has been derived for the drift transistor without relating it to a specific circuit. We have obtained

$$U = \frac{|\alpha|^2}{4r_b'\,C_c^2\omega^2\left[r_e - \frac{\mathrm{Im}\,(\alpha)}{\omega\,C_c} @\ \omega\right]} \qquad (5)$$

Gains with negative sign are predicted, and this means that a negative resistance is present at one of the ports of the network. Values of negative resistance are apparent in equation (5), where in a region of the positive imaginary part of α, a resistance with a negative sign will appear, e.g., when the phase angle is in a range of $180^\circ < (\omega) < 360^\circ$, and provided that $\mathrm{Im}(\alpha)/\omega\, C_c > r_e$. Straightforward circuit analysis of the impedance, Z_{eb},

$$\alpha_i = \alpha_o \frac{\exp\left\{-jm \, f/f_c\right\}}{1 + j \, f/f_c}$$

for $\alpha_o = .985$

$m = 1.1$

$f_c = 600$ MC

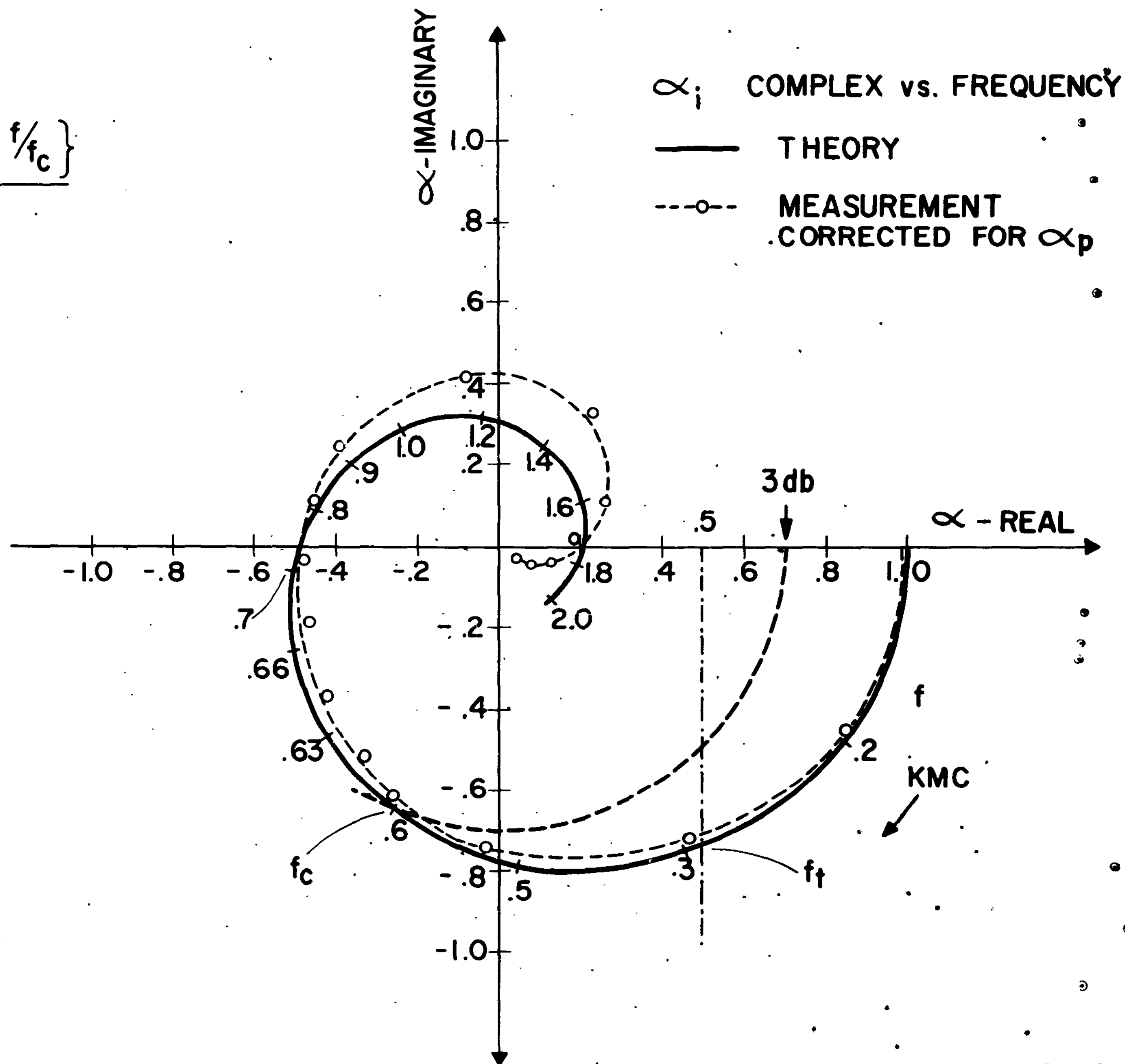


FIG. 8

looking into the emitter-to-base terminals of a transistor, when the collector output is terminated into an inductive reactance, L_c, (See Figure 9) leads to

$$Z_{eb} = r_e + r_b' \left[\frac{(1 - a - \omega^2 L_c C_c)(1 \quad \omega^2 L_c C_c) - b\omega r_b' C_c}{(1 - \omega^2 L_c C_c)^2 + (\omega r_b' C_c)^2} \right] \quad (6)$$

$$- jr_b' \left[\frac{b(1 - \omega^2 L_c C_c) + \omega r_b' C_c (1 - a - \omega^2 L_c C_c)}{(1 - \omega^2 L_c C_c)^2 + (\omega r_b' C_c)^2} \right]$$

where a is the real part and b is the imaginary part of the complex α.

If L_c is tuned such that at the frequency

$$\omega_o = \frac{1}{L_c C_c} \quad (7)$$

the imaginary part is a positive maximum (phase angle approximately 270°) e.g., coaxial line element; then

$$Z_{eb} = \left[r_e - \frac{b}{\omega_o C_c} \right] + j \left[\frac{a}{\omega_o C_c} \right] \quad (8)$$

The real part of this impedance is exactly the term which appears in the denominator of equation (5) and can become negative when $b/\omega_o C_c > r_e$. By proper termination at the input terminals then, a negative resistance oscillator will result at ω_o. (Figure 9)

To utilize the negative resistance between the collector-to-emitter terminal as derived by the equation

$$Z_{ce} \simeq \frac{1 - \alpha}{\omega C_c} = \frac{1}{\omega C_c} [|\alpha| \sin \quad - (|\alpha| \cos \quad - 1)] \quad (9)$$

LETTER PROGRESS AND INTERIM ENGINEERING REPORT

For

HIGH FREQUENCY TRANSISTOR STUDY

This report covers the period 10 March through 30 June 1961

Advanced Development Laboratory
Lenkurt Electric Company
San Carlos, California

For

Hughes Semiconductors
(Purchase Order Number 1-877651-M11A)

July 28, 1961

PART III

INTRODUCTION

This document constitutes the letter progress report for the period of June 30 through July 31 and the first interim engineering report on this project.

In this study, a Philco coaxial transistor, L-5431, has been used with a coaxial cavity to amplify signals at frequencies ranging from 1-2KMC and 2-4KMC. Two cavities were designed, one for each frequency range; Coaxial Cavity I, illustrated in Figure 1, for the 1-2KMC frequency range, and Coaxial Cavity II, Illustrated in Figure 2, for the 2-4KMC frequency range. At this time, promising results have been obtained with Coaxial Cavity I only, although no overall amplification of the signal between the signal generator and the output of this cavity has been obtained. The gain of the transistor alone has been determined, however, by measuring the difference in the insertion losses of the system, first with the transistor power supply on, and then with the transistor power supply off.

AMPLIFIER CIRCUIT COMPONENTS

In the amplifier circuit, illustrated in Figure 3, coaxial capacitor 3 acts as a variable blocking capacitor which prevents the dc bias of the emitter junction from reaching the signal generator. This capacitor was built by replacing the resistor element of an attenuator pad with a cylindrical trimmer (3-10pf). RF chokes 9 and 18 consist of several turns of hook-up wire. These open wires, as well as the lead coming from T-section 4, are responsible for some of the overall losses within the system. Capacitors 10 and 19, each with a capacitance of approximately 1500pf, serve as RF bypass capacitors and also protect the transistor from switching transients. Potentiometers 11 (500 ohms) and 12 (10,000 ohms) are used for fine and coarse adjustments, respectively, of the emitter current. Fixed resistors 13 (200 ohms) and 17 (100 ohms) protect the transistor. In this amplifier circuit, the emitter currents range from 1 to 4ma at points of resonance.

CONSTRUCTION OF COAXIAL CAVITY I

The construction of Coaxial Cavity I is shown in Figure 4 which has been drawn on an approximate 1:1 scale.

PERFORMANCE OF COAXIAL CAVITY I

Figure 5 shows the voltage gain of the signal generator and the output of Coaxial Cavity I with the power to the transistor first on, and then off. Since the RF meter used for these measurements gives comparative voltage readings, the difference between the insertion losses of the circuit with the power to the transistor on and then off is a measure of the transistor amplification when the SWR of the line is 1.

As may be seen from Figure 6, the amplification or voltage gain of the transistor exceeds 30 db at 1050MC. This gain is a result of the change in impedance at a resonant condition (although the transistor is nearly unstable at a resonant condition), but the gain decreases to 7 db at a frequency of 2KMC. It may also be noted that the cavity is multi-resonant. Since these resonances occur after adjusting the length of the cavity and the emitter current, their presence is not objectionable.

Figure 7 indicates the performance of Coaxial Cavity I. Thus, this cavity has a Q_L of 122 at 1KMC and a Q_L of 445 at 2KMC. These low values for the loaded Q's are the "lossy" contact between the inner and outer tube of the cavity as well as losses in the magnetic output pickup loop installed at the front end of the cavity.

CONCLUSION

1) The optimum performance of Coaxial Cavity I occurs at 2KMC, and maximum transistor amplification occurs at 1KMC.

2) The amplifier circuit may be improved by decreasing the insertion loss of the circuit components and increasing the Q of both

Coaxial Cavity I and Coaxial Cavity II. Thus, the circuit will be improved by:

a) Replacing RF chokes 9 and 18 with coaxial chokes.

b) Changing the output pick-up loop to a capacitative type of pick-up loop.

c) Using cylindrical cavities (without center stubs) for amplification.

d) Improving the design of the miniature 2-4KMC cavity, Coaxial Cavity II.

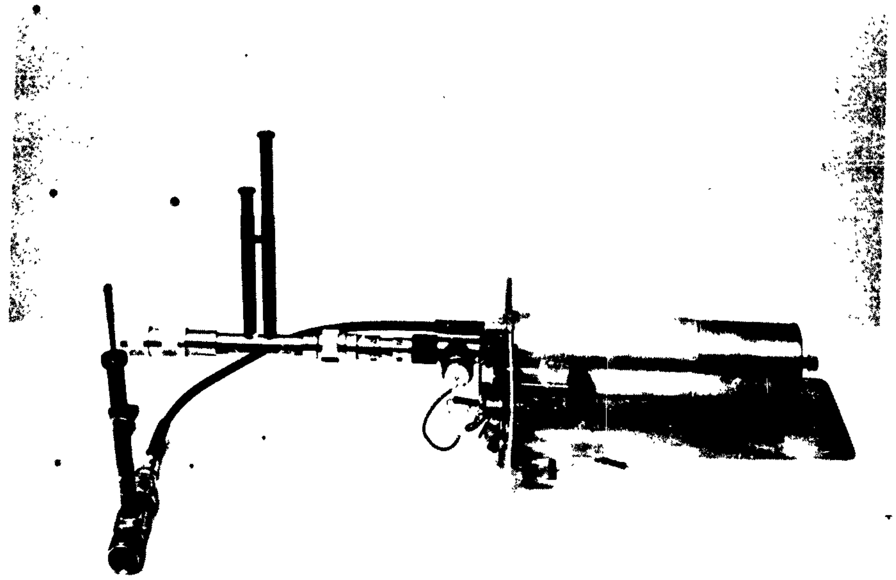


FIGURE 1. COAXIAL CAVITY I

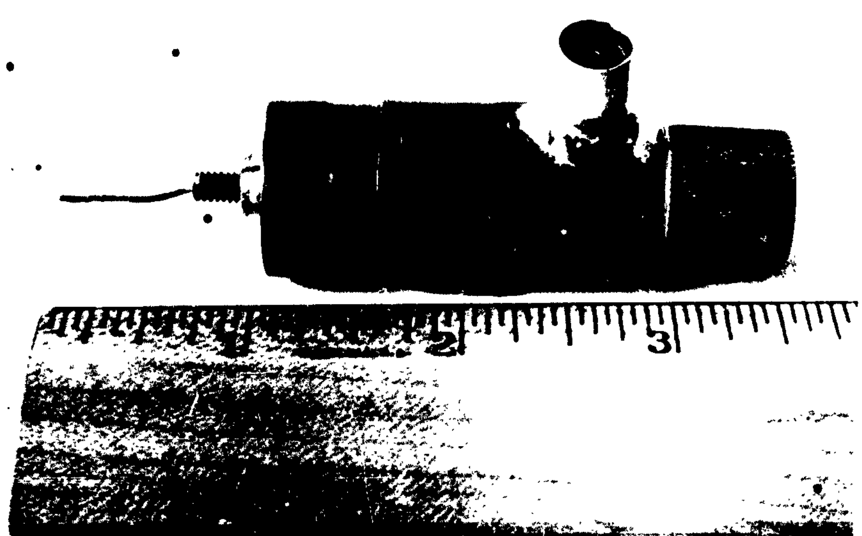
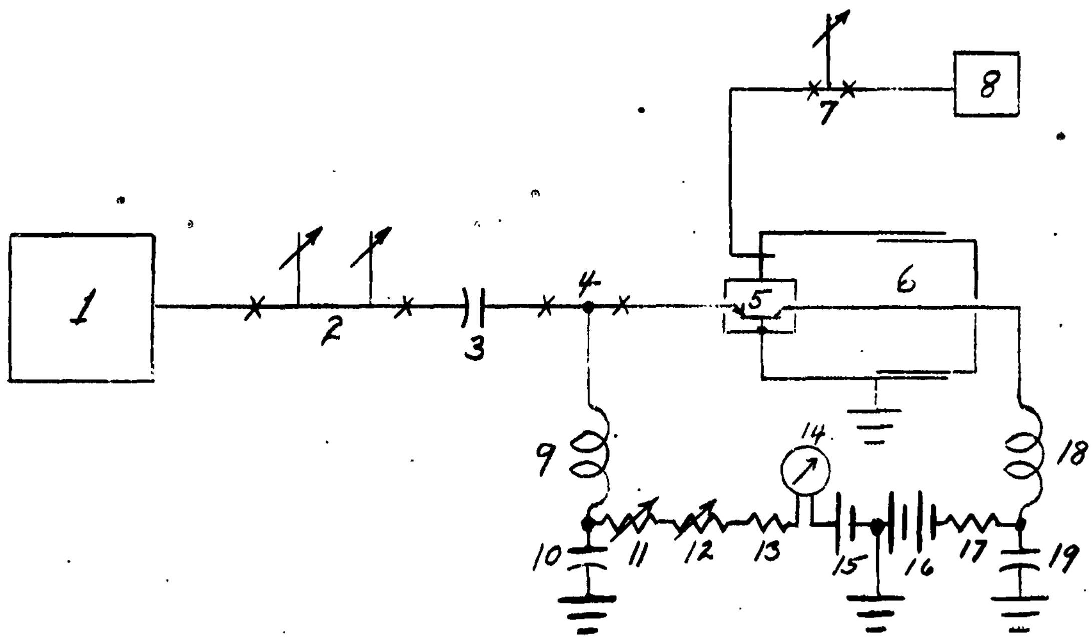
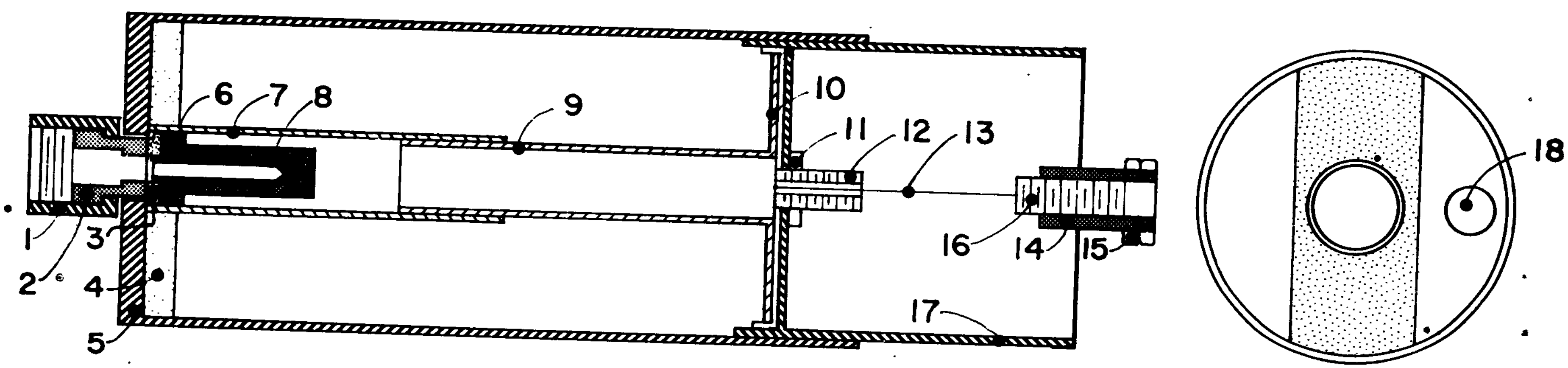


FIGURE 2. COAXIAL CAVITY II

Figure 3. Schematic Diagram of the Amplifier Circuit



1. 614A HP Signal Generator
2. Double stub tuner
3. Adjustable coaxial blocking capacitor
4. T junction
5. Coaxial transistor (Philco L-5431)
6. Coaxial cavity
7. Single stub tuner
8. 91 CA Boonton RF voltmeter
9. Choke coil
10. By-pass capacitor
11. Resistor pot (fine adjustment)
12. Resistor pot (rough adjustment)
13. Resistor
14. Milliammeter
15. Emitter bias cell, 3 volts
16. Collector bias cell, 7.5 volts
17. Resistor
18. Choke coil
19. By-pass capacitor



Inside view of outer tube

1. Part of "N" connector
2. Bushing
3. Insulator
4. Insulator
5. Outer tube
6. Chuck holder
7. Center conductor of outer tube
8. Chuck
9. Center conductor of inner tube
10. 1mm mylar insulator
11. Nut
12. Nylon screw
13. Heavy lead wire
14. Nylon bolt
15. Nut
16. Screw
17. Inner tube
18. Hole for coupling loop

Figure 4. Coaxial Cavity I

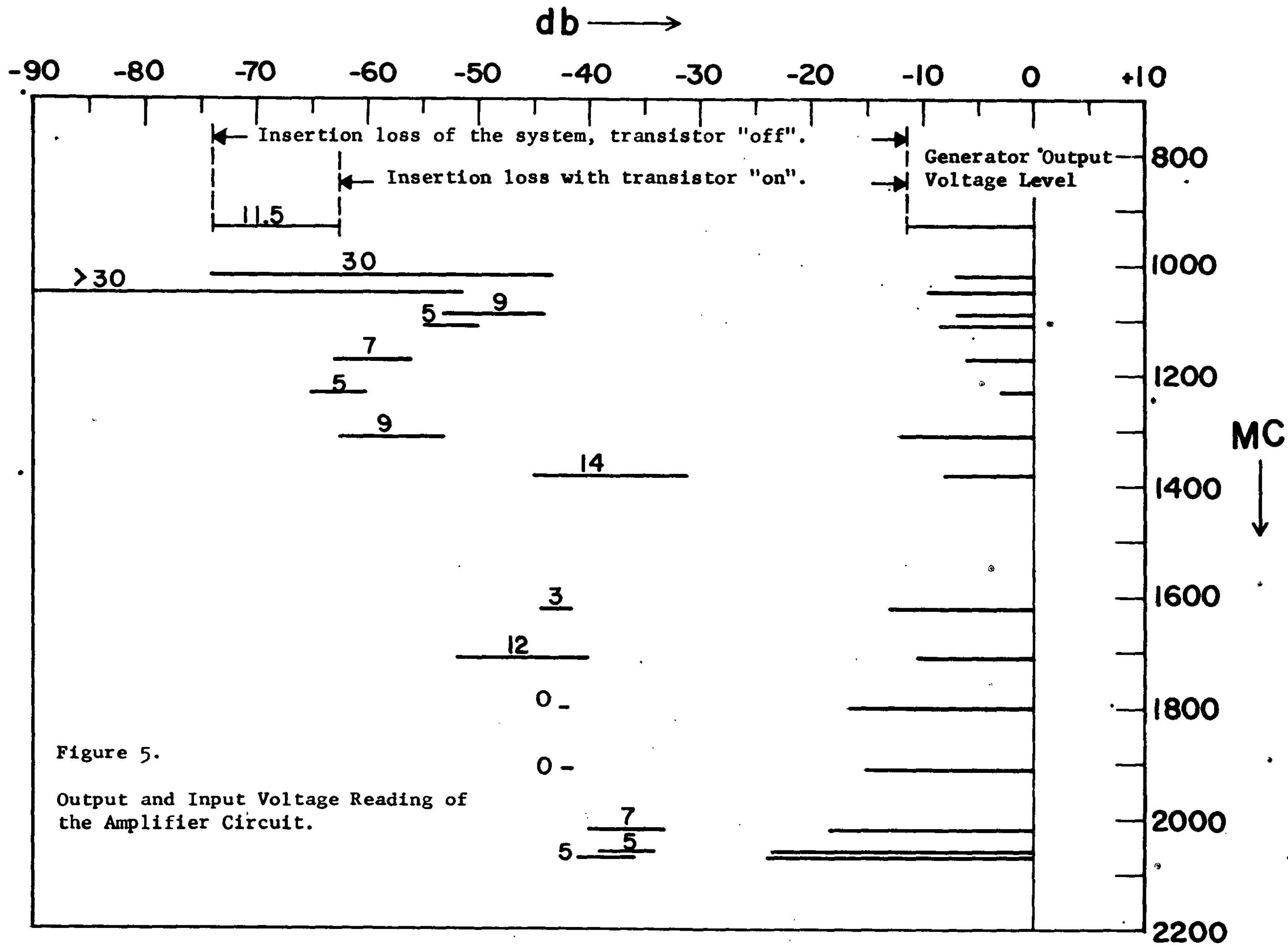


Figure 5.

Output and Input Voltage Reading of the Amplifier Circuit.

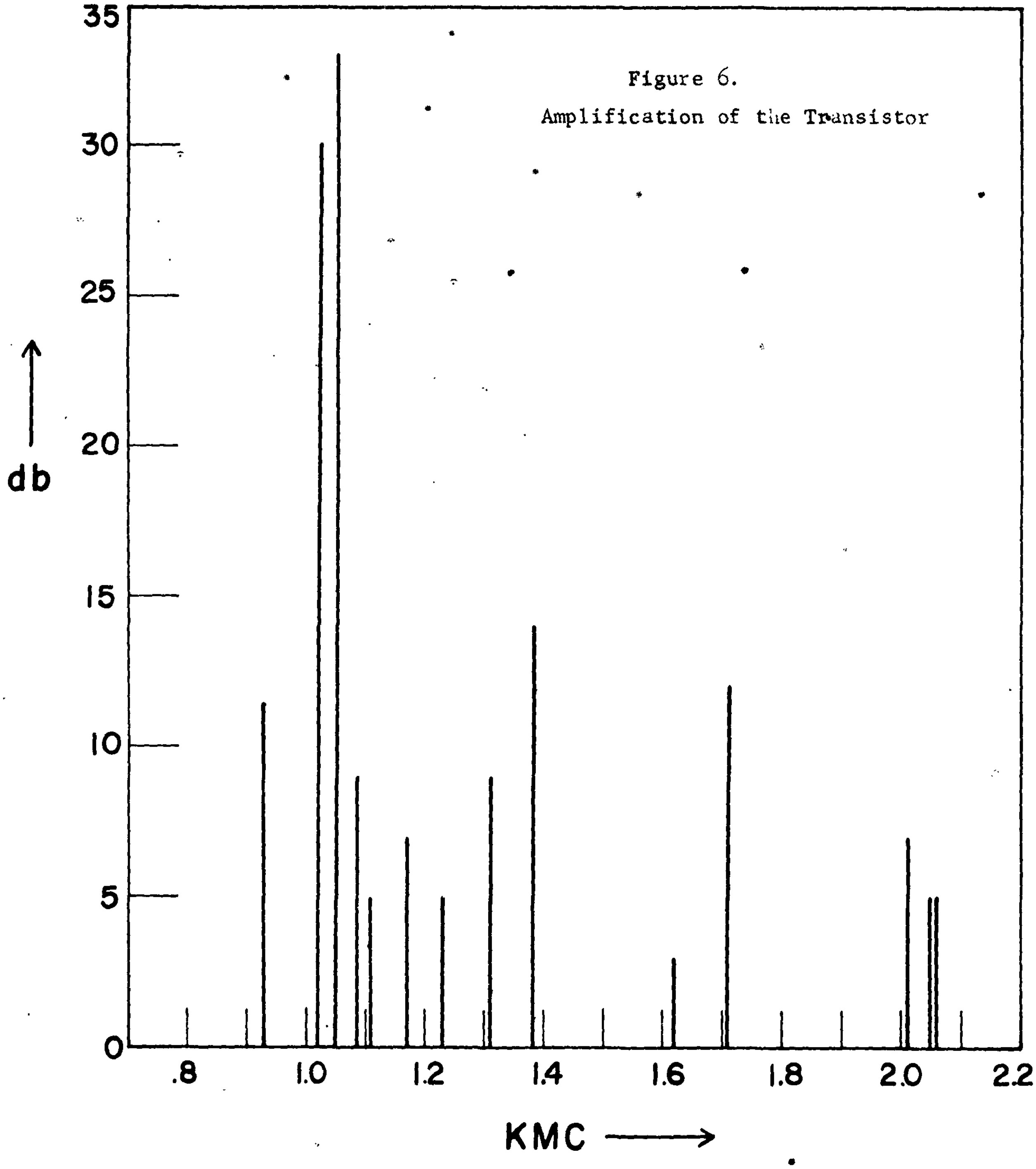


Figure 6.
Amplification of the Transistor
35
30
25
20
15
10
5
0
db
.8
1.0
1.2
1.4
1.6
1.8
2.0
2.2
KMC

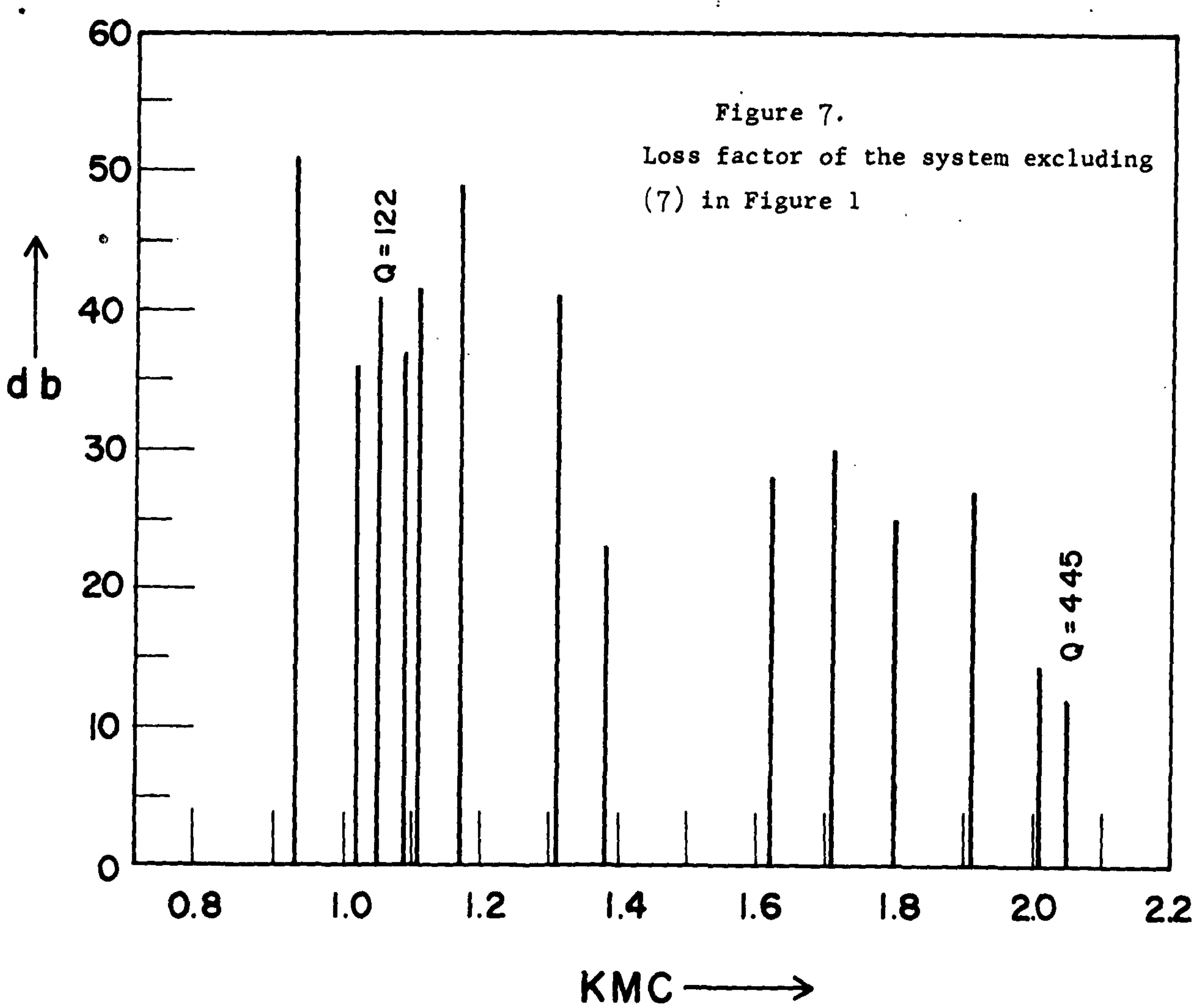


Figure 7.
Loss factor of the system excluding (7) in Figure 1
60
50
40
30
20
10
0
d b
Q = 122
Q = 4.45
0.8
1.0
1.2
1.4
1.6
1.8
2.0
2.2
KMC